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**1229; VAN GINDERDEUREN E. A.:**  
**'CORDIC-BASED HIFI DIGITAL FM**  
**DEMODULATOR ALGORITHM FOR COMPACT**  
**VLSI'**

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## Description

The invention relates to a receiver comprising an A/D converter for digitally sampling an analog signal modulated on a carrier frequency at a first sampling frequency, consecutively coupled to a digital quadrature mixer stage for a carrier frequency shift of the digitized modulated signal from the A/D converter, provided with first and second signal outputs *via* which the digital quadrature mixer stage supplies a pair of carrier frequency converted phase quadrature signals, a digital filter device for selecting the phase quadrature signals of the quadrature mixer stage and for decimating the sampling frequency from said first sampling frequency to a second sampling frequency, and a digital demodulation device.

A receiver of this type is known from European Patent Application EP-A-35166.

The known receiver is of the direct conversion type in which the analog RF reception signals are first digitized at the first sampling frequency and subsequently multiplied in a pair of multiplier circuits of the digital quadrature mixer stage by a pair of tunable digital phase quadrature oscillator signals which are supplied from a digital quadrature tuning oscillator. The digitised RF reception signals are thereby split in phase into I (in-phase) and Q (quadrature) signals which differ mutually 90° in phase, and the digitized RF reception signal whose RF carrier frequency corresponds to the tuning frequency of the digital phase quadrature oscillator signals is also shifted in frequency towards the baseband. The desired digital I and Q baseband phase quadrature signals are selected in the digital filter device in which simultaneously the sampling frequency is reduced (decimated) from the first sampling frequency to the second sampling frequency. The digital I and Q baseband signals thus selected are subsequently demodulated in the digital demodulation device to a single digital baseband signal which is applied to a reproducing device *via* a D/A converter.

For an effective realisation of the known receiver the use of complex digital multipliers in the oscillator circuit as well as in the mixer stage is inevitable. An integrated embodiment therefore requires a comparatively large crystal surface. Moreover, the quadrature tuning oscillator should generate tunable digital sine and cosine oscillator signals having a comparatively low distortion. Particularly in the frequency range of normal radio and television broadcast transmitters it is difficult to comply with this distortion requirement, which makes the known receiver less suitable for use as a receiver of broadcast signals.

The invention has for its object to provide a receiver which is easier to realise as compared with said known digital receiver and which in an integrated form requires less crystal surface, while a correctly chosen dimensioning renders it suitable, *inter alia*, for receiving RF radio or TV broadcast signals.

This object is solved by a receiver comprising an A/D converter for digitally sampling an analog signal modulated on a carrier frequency at a first sampling frequency, consecutively coupled to a digital quadrature mixer stage for a carrier frequency shift of the digitized modulated signal from the A/D converter, the digital quadrature mixer stage being provided with first and second signal outputs *via* which it supplies a pair of carrier frequency converted phase quadrature signals, a digital filter device for selecting the phase quadrature signals of the quadrature mixer stage and for decimating the sampling frequency from said first sampling frequency to a second sampling frequency, and a digital demodulation device being coupled to the digital filter device, which is characterized in that the digital quadrature mixer stage comprises a first coordinate rotation digital computer (Cordic) in the rotation mode, having at least a first signal input which is coupled to an output of the A/D converter and a phase signal input which is coupled to an output of a digital sawtooth generator, which digital sawtooth generator supplies a periodical digital sawtooth-shaped phase signal to the phase signal input of said computer for a periodical  $2\pi$  phase rotation of at least the signal applied to the computer *via* the first signal input at a repetition frequency which is equal to the magnitude of said carrier frequency shift, said computer supplying at the first and second signal outputs said pair of carrier frequency converted phase quadrature signals to the digital filter device.

The use of a coordinate rotation digital computer, hereinafter referred to as Cordic for short for fast digital trigonometric computations is known from the article "The Cordic Trigonometric Computing Technique", published in "IRE Transactions on Electronic Computers", September 1959 by J.E. Volder. The computations are effected *via* simple signal processing operations such as binary shifts, additions, subtractions and calling prestored constants. The Cordic thus has a very simple and compact integrable circuit structure which in an integrated form requires a comparatively small crystal surface.

The invention is based on the recognition that a Cordic, which operates in the so-called rotation mode, in which mode a polar signal representation is converted into a cartesian representation, can function as a mixer oscillator circuit in combination with a sawtooth generator at the phase signal input in the case of a correctly chosen dimensioning. The carrier frequency of the input signal of the Cordic is converted or shifted over a frequency which is equal to the repetition frequency of the signal supplied by the sawtooth generator to the first Cordic, which signal will hereinafter be referred to as phase signal for short.

When using the measure according to the invention, the digital mixer oscillator circuit constituted by the combination of the first Cordic in the rotation mode and the sawtooth generator effects a conversion of the digitized signals modulated on said carrier frequency, dependent on the repetition frequency of the phase signal to the baseband or to a fixed

intermediate frequency without the formation of digital sine and/or cosine oscillator signals and/or accurate signal multiplications being required.

This provides a simple integrable realisation of a digital receiver which is suitable for receiving RF signals in a very large frequency range, particularly RF radio and television broadcast signals.

In said receiver according to the invention the combination of the first Cordic and the sawtooth generator can be used as part of a demodulation circuit to which an intermediate frequency (IF) signal at a fixed IF carrier frequency is applied, for example, from a tuning circuit. The repetition frequency of the sawtooth generator is then to be chosen at a suitable fixed value, for example, a value which is equal to the last-mentioned IF carrier frequency.

However, it is alternatively possible to use the combination constituted by the first Cordic and the sawtooth generator as an RF tuning circuit. A preferred embodiment of a receiver according to the invention is therefore characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal is variable in a range which corresponds to the frequency reception range of the receiver.

For a direct conversion of a desired RF reception signal to the baseband, such a so-called direct conversion receiver according to the invention is characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal of the digital sawtooth generator is equal to said carrier frequency of the analog signal applied to the A/D converter for a direct frequency conversion to the baseband in the computer.

For a conversion of a desired RF reception signal to a fixed IF, such a so-called superheterodyne receiver according to the invention is characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal deviates by a fixed intermediate frequency value from said carrier frequency of the analog signal applied to the A/D converter.

A receiver according to the invention is preferably characterized in that the phase rotation caused by the periodical digital sawtooth-shaped phase signal varies linearly over a range of  $2\pi$  radials in each period. The occurrence of non-linear distortion is thereby inhibited.

An improved noise behaviour is achieved in a further preferred embodiment of a receiver according to the invention which is characterized in that the first sampling frequency is a non-integral number of times the carrier frequency of the analog signal applied to the A/D converter, which number is larger than two.

A further preferred embodiment of a receiver according to the invention is characterized in that the digital demodulation device comprises a second coordinate rotation digital computer (CORDIC) in the vectoring mode, having first and second signal inputs to which the selected phase quadrature signals of the digital filter device at the second sampling frequency are applied, and an output which is coupled to a signal reproducing device.

When using this measure, a signal processing operation is performed in the second coordinate rotation digital computer, hereinafter referred to as second Cordic for short, which is complementary with respect to that in the first Cordic, and a correct demodulation/conversion is obtained of the selected and sampling frequency-decimated phase quadrature baseband signals into a single-phase signal which comprises the desired baseband information.

In the case of FM RF signals the last-mentioned single-phase signal becomes available at the phase signal output of the second Cordic and only a signal differentiation is required to obtain the baseband modulation signal. To this end an FM receiver according to the invention is characterized in that said output of the second coordinate rotation digital computer (CORDIC) is the phase signal output and is coupled to the signal reproducing device via a differentiating circuit.

An embodiment of the receiver according to the invention which is suitable for receiving AM RF signals such as, for example, AM radio or AM-VSB TV signals is preferably characterized in that said output is a first signal output of the second coordinate rotation digital computer (Cordic) to which an output signal is supplied which varies with the amplitude of the input signals of said second computer, said first signal output being coupled to the signal reproducing device via an amplitude detector.

The first Cordic makes the digital baseband signal in the form of a baseband phase quadrature I and Q signal pair at the first sampling frequency available at said first and second signal outputs.

In order to simplify selection of the I and Q baseband signals, a preferred embodiment of a receiver according to the invention is characterized in that the first coordinate rotation digital computer has a second signal input, while a Hilbert signal transformation is effected between the A/D converter and one of the two signal inputs and a signal delay compensation is effected between the A/D converter and the other signal input.

To realise a correct phase quadrature relationship between the signals at the two signal inputs of the first Cordic, a further preferred embodiment is characterized in that a digital anti-symmetrical finite impulse response filter is incorporated between the A/D converter on the one hand and the first and second signal inputs of the first coordinate rotation digital computer on the other hand, which filter comprises a series arrangement of  $2n+1$  ( $n = 0, 1, \dots$ ) serial pairs of delay circuits, the output of each pair of delay circuits as well as the input of the series circuit being coupled to an adder circuit via a weighting factor multiplier circuit, the common connections between the delay circuit of the  $(n+1)^{\text{th}}$  pair being coupled to the first signal input and the output of the adder circuit being coupled to the second signal input of the first coordinate rotation digital computer.

A digital filter device which can be realised very easily for selecting and decimating the output signals of the first Cordic is characterized in that said digital filter device has a sub-sample filter section which comprises a cascade circuit of an  $n^{\text{th}}$  order comb filter and an accumulate and dump circuit comprising an adder circuit having first and second signal inputs and a signal output, said first signal input being coupled to an output of the comb filter device and said signal output being coupled to the second signal input via a delay circuit for a signal delay of an input sampling frequency over one period, the output signal of the adder circuit being sampled at an output sampling frequency which is  $1/2n$  the input sampling frequency and the accumulate and dump circuit being reset after each last-mentioned sampling.

The invention will be described in greater detail with reference to the Figures shown in the accompanying drawings which are only given by way of example and in which corresponding elements have the same references.

In the drawings

Fig. 1 shows a functional block diagram of a receiver according to the invention of the direct conversion type.

Fig. 2 shows a functional block diagram of a sawtooth generator in combination with a Cordic in the rotation mode for use in a block diagram of the receiver of Fig. 1.

Fig. 3 shows the time-dependent variation of the sawtooth-shaped phase signal of the sawtooth generator in the receiver of Fig. 1 and/or 2.

Fig. 4 shows a practical embodiment of a part of the first Cordic in the rotation mode and the sawtooth generator in the circuit of Fig. 2.

Fig. 5 shows a block diagram of a Hilbert transformation circuit.

Fig. 6 shows a block diagram of a digital sub-sample filter for use in the digital baseband quadrature filter device of the receiver of Fig. 1.

Fig. 1 shows a receiver according to the invention of the direct conversion type, with an antenna input 1 for connecting a radio frequency (RF) antenna device A thereto to which there are consecutively coupled: a first RF input filter 2, a gain-controlled RF amplifier device 3, a second RF input filter 4, an A/D converter 5, a digital RF filter 6 for converting the RF signals digitized in the A/D converter 5 into digital in-phase (I) and quadrature (Q) RF signals, a first coordinate digital computer (Cordic) 9 in the rotation mode having first and second signal inputs 16 and 17 and a phase signal input 18, first and second signal outputs 19 and 20, and a phase signal output 21, a digital baseband quadrature filter device 10, 11, a second Cordic 12 in the vectoring mode having first and second signal inputs 22 and 23 and a phase signal input 24 and first and second signal outputs 25 and 26 and a phase signal output 27.

The receiver shown is an FM receiver and to this end it comprises a signal differentiating circuit 13 which is coupled to the phase signal output 27 of the second Cordic, which signal differentiating circuit 13 is connected to a signal processing device 14 for further baseband processing and reproduction.

The first RF filter device 2, the gain-controlled RF amplifier device 3 and the second RF filter device 4 jointly constitute an analog RF input section in which amplitude variations of the RF antenna input signal are adapted to the dynamic range of the A/D converter 5 and in which the RF reception range of the receiver is selected. The gain control signal for the RF gain-controlled amplifier device 3 is supplied by the A/D converter 5. In the A/D converter 5 the RF signals in said RF reception range of the receiver are digitized in a first sampling frequency which should be at least twice the highest frequency in the RF reception range. In a practical embodiment of the receiver shown with an RF reception range between 87.5 MHz and 108 MHz, the first sampling frequency was 350 MHz. The RF reception signals thus digitized are subsequently applied to the digital RF filter 6 for conversion into a pair of phase quadrature RF signals  $x_i$  and  $y_i$  which are applied to the first and second signal inputs 16 and 17, respectively, of the first Cordic 9. A digital anti-symmetrical FIR (Finite Impulse Response) filter which is known *per se* from, for example, the book "Discrete Time Signal Processing" by A.W.M. van den Enden and N.A.M. Verhoeckx, published by Prentice Hall International (UK) Ltd. in 1989, paragraph 8.2.4, pp. 208-211 can be used as a digital RF filter 6.

The operation of the first Cordic 9 is known *per se*, for example, from the article "A unified algorithm for elementary functions" by J.S. Walther published in "Spring Joint Computer Conference, 1971", pp. 379 to 385. As is known from this article, a Cordic computes in the rotation mode, like the first Cordic 9, the cartesian coordinates of a signal vector  $x_0$  and  $y_0$  obtained by rotating an input signal vector formed by the input phase quadrature signals  $x_i$  and  $y_i$  at the first and second signal inputs 16 and 17 of the first Cordic 9 through a phase angle  $Z_0$  which is applied to the first phase signal input 18 of this first Cordic 9. Since the first Cordic 9 is operative in the rotation mode, a fixed signal value is supplied to the phase signal output 21, which value is equal to 0 or, within a given tolerance deviation, is equal to 0 and which will hereinafter be briefly referred to as the zero value. Each of the circuits 2-13 mentioned so far is known *per se* and, apart from their respective functions, need not be further described to understand the invention.

According to the invention, the receiver shown comprises a digital sawtooth generator SG which is coupled to the phase signal input 18 of the first Cordic 9. A digital tuning data  $f_t$  is applied to the digital Sawtooth generator SG via a tuning signal input T.

The digital sawtooth generator SG supplies a digital phase angle value  $Z_0$  to the phase signal input 18 of the first

Cordic 9, which value varies periodically in such a way that the successive digital values of the phase angle  $Z_0$  can be considered to be successive signal samplings of a periodical digital sawtooth-shaped phase signal  $z$  having a phase rotation of  $2\pi$  rad. per period. In other words, the phase angle value  $Z_0$  varies in each period of the phase signal  $z$  while it monotonically increases or decreases over a range of  $2\pi$  radial. The repetition frequency of the phase signal  $z$  is adjusted at a desired value by means of the digital tuning data  $f_t$ . The first Cordic 9 consequently operates as a digital quadrature mixer stage in which a frequency conversion or shift is effected over a frequency which is equal to the repetition frequency of said periodical digital sawtooth-shaped phase signal. In a receiver of the direct conversion type the repetition frequency of the phase signal is chosen to be equal to the RF carrier frequency of the desired RF reception signal. The output signals  $x_0$  and  $y_0$  at the first and second signal outputs 19 and 20 of the first Cordic 9 thus constitute a pair of baseband phase quadrature signals.

As already mentioned hereinbefore, the variable digital tuning data  $f_t$ , with which the repetition frequency of the sawtooth-shaped phase signal  $z$  can be varied over a desired frequency range, is applied to the sawtooth generator SG via a tuning control input T. For a direct conversion in the first Cordic 9 the variation range of the repetition frequency of the phase signal  $z$  should correspond to the RF reception range of the receiver. The baseband phase quadrature signal  $x_0, y_0$  is subsequently applied to the digital baseband phase quadrature filter device 10, 11 in which the baseband phase quadrature signals  $x_0$  and  $y_0$  are selected and in which the sampling frequency is decimated or decreased from said first value  $fs_1$  to a second value  $fs_2$  resulting in a baseband phase quadrature signal  $x'_i, y'_i$ . Such a digital baseband phase quadrature filter device 10, 11 is known *per se*, for example, from European Patent Application EP 35166 and includes a selective low-pass filter which is operative for the two baseband phase quadrature signals  $x_0$  and  $y_0$ , as well as a decimating or sub-sample low-pass filter for decimating the sampling frequency from  $fs_1$  to  $fs_2$ .

If the receiver is an FM broadcast receiver, a sub-sample filter as is shown in Fig. 6 is preferably used for each one of the baseband phase quadrature signals  $x_0$  and  $y_0$ . In a practical embodiment of such an FM broadcast receiver the last-mentioned sub-sample filter is succeeded by a wave digital filter (not shown) as is known *per se* from the article "Wave digital filters: Theory and Practice" by A. Fettweis, Proceedings of the IEEE, vol. 74, no. 2, February 1986 in which a further selection and decimation takes place.

If the receiver is a TV receiver, an asymmetrical polyphase filter as is known *per se* from United States Patent no. 4,914,408 is preferably used in combination with a sub-sample filter adapted thereto for said selective low-pass filter.

The baseband phase quadrature signal  $x'_i, y'_i$  thus obtained in the second sampling frequency  $fs_2$  is applied to the first and second signal inputs 22 and 23 of the second Cordic 12 which operates in the vectoring mode. As is known from the above-mentioned publication by Walther, a conversion from a cartesian signal representation to a polar signal representation takes place in the second Cordic 12 because it operates in the vectoring mode, and it is possible to determine the angle and angle variations of an input signal vector such as, for example,  $x'_i, y'_i$  with respect to a fixed reference coordinate, for example, the X axis. In this vectoring mode the phase signal input 24 of the second Cordic 12 conveys the zero value and a zero value level is supplied to the second signal output 26 of the Cordic 12. In the case of an FM receiver the angle modulation of the FM reception signal is obtained at the phase signal output 27 of the second Cordic and this angle should be differentiated. This is effected in the signal differentiating circuit 13 at whose output the desired FM modulation signal in the baseband becomes available. Subsequently, a further signal processing such as, for example, a D/A conversion and reproduction is effected in the signal processing device 14.

As can be deduced from the above-mentioned publication by Walther, the amplitude or absolute magnitude of the input signal vector  $x_3, y_3$  applied to the first and second signal inputs of the second Cordic 12 is obtained at the first signal output 25 of this Cordic. Consequently it is possible to render the receiver shown easily suitable for receiving amplitude-modulated (AM) RF reception signals by coupling the first signal output 25 of the second Cordic 12 to a baseband signal processing and reproducing circuit 15 or by connecting it to the signal processing device 14 via a switch (not shown).

It is alternatively possible to use the quadrature mixer stage constituted by the first Cordic 9 in combination with the sawtooth generator SG for a fixed frequency conversion of an intermediate frequency (IF) signal of a superheterodyne receiver. To this end such a superheterodyne receiver (not shown) may comprise an analog tuning circuit which is incorporated between the RF input section 2, 3, 4 on the one hand and the A/D converter 5 on the other hand, or it may comprise a digital quadrature tuning circuit which is known from said European Patent Application EP-A-35166, which circuit replaces or precedes the digital RF filter 6. The repetition frequency of the sawtooth-shaped phase signal  $z$  supplied by the sawtooth generator SG should then correspond to the sum or the difference of the desired RF carrier frequency and the intermediate frequency of the superheterodyne receiver. It stands to reason that the intermediate frequency can be chosen at a value which is above the RF reception frequency range (for example, in the case of a double superheterodyne receiver in which a high first intermediate frequency and a lower second intermediate frequency are used) or which is below this range.

It is to be noted that the phase split, which is realised in the digital RF filter 6 to obtain a pair of phase quadrature signals, is not essential for the use of the invention. The supply of phase quadrature signals  $x_i, y_i$  to the signal inputs 16 and 17 of the first Cordic 9 prevents the occurrence of mixed sum components, or mixing products at the signal

outputs 19 and 20, which are located at the sum frequency of the carrier frequency of the last-mentioned phase quadrature signals  $x_i$ ,  $y_i$  and the repetition frequency of the phase signal  $z$ . Consequently, the selection of the desired mixed difference components in the digital baseband phase quadrature filter 10, 11 is simplified to a considerable extent.

The digital RF filter 6 can be dispensed with by imposing stricter selectivity requirements on the last-mentioned baseband phase quadrature filter 10, 11. The digitized RF reception signal of the output of the A/D converter 5 can then be applied directly to one of the two signal inputs 16 and 17, while the other signal input is connected to a zero value level.

Fig. 2 shows a functional block diagram of the digital quadrature mixer stage constituted by the first Cordic 9 and the sawtooth generator SG according to the invention. In the version shown this stage comprises a cascade circuit of eleven iteration sections IS1 to IS11 in which one of eleven consecutive iterative angle convergence steps is performed. As is known from the above-mentioned article by Walther, the associated changes in the signal values of the signal or signals applied to the signal input 16 and/or the signal input 17 are defined for each iteration step in the sections IS2 to IS11 by the following equations:

$$X(j+2) = X(j+1) - \text{sign } Z(j+1) \cdot 2^{-j} \cdot Y(j+1)$$

$$X(j+2) = Y(j+1) + \text{sign } Z(j+1) \cdot 2^{-j} \cdot X(j+1)$$

$$Z(j+2) = Z(j+1) - \text{sign } Z(j+1) \cdot \arctan 2^{-j}$$

with  $j = 0 \dots 9$  for a given choice of the angle convergence.

For the section IS1 it holds that:

$$X1 = -\text{sign } Z_0 \cdot Y_0$$

$$Y1 = \text{sign } Z_0 \cdot X_0$$

$$Z1 = Z_0 - \text{sign } Z_0 \cdot \pi/2$$

The references  $X_0$  and  $Y_0$  denote the sampling values of  $x_i$  and  $y_i$ , respectively, and  $\text{sign } Z_0$  denotes the sign of the sampling value  $Z_0$  of  $z$ .

In a so-called  $z$  branch of the sections IS1 to IS11, which will hereinafter be described in greater detail, the input angle value  $Z_0$  is rotated iteratively via a series of fixed converging angles (see Table I below) towards zero or at least towards a residual value deviating therefrom within the tolerance deviation. For each iteration section the sign, or the direction of the fixed rotation angle is defined with which the desired angle convergence is obtained. In the first iteration section IS1  $\pi/2$  is added to  $Z_0$  or subtracted from  $Z_0$  in dependence on the sign of  $Z_0$ , resulting in a new angle value  $Z1$ .

In the second iteration section IS2  $\pi/4$  is added to  $Z1$  or subtracted from  $Z1$  in dependence on the sign of  $Z1$ , resulting in  $Z2$  such that  $Z2 < Z1$ , etc., etc.

The Table below shows through which angles the input angle value  $Z_0$  is successively rotated in the consecutive iteration sections. It is based on a 12-bit representation in so-called two's complement mode, in which  $2^{12} = 4096$  corresponds to  $2\pi$ .

Table I:

| $\phi$ | RAD                    | RAD         | DEG  | DEC  | + BIN        | - BIN        |
|--------|------------------------|-------------|------|------|--------------|--------------|
| 1      | $2 \cdot \arctan(2^0)$ | 1.570796327 | 90   | 1024 | 010000000000 | 110000000000 |
| 2      | $\arctan(2^0)$         | 0.785398163 | 45   | 512  | 001000000000 | 111000000000 |
| 3      | $\arctan(2^{-1})$      | 0.463647609 | 26.6 | 302  | 000100101110 | 111011010010 |
| 4      | $\arctan(2^{-2})$      | 0.244978663 | 14.0 | 160  | 000010100000 | 111101100000 |
| 5      | $\arctan(2^{-3})$      | 0.124354995 | 7.1  | 81   | 000001010001 | 111110101111 |
| 6      | $\arctan(2^{-4})$      | 0.062418810 | 3.6  | 41   | 000000101001 | 111111010111 |

Table I: (continued)

| $\phi$ | RAD                    | RAD         | DEG | DEC | + BIN        | - BIN        |
|--------|------------------------|-------------|-----|-----|--------------|--------------|
| 7      | $\arctan(2^{\wedge}5)$ | 0.031239833 | 1.8 | 20  | 000000010100 | 111111101100 |
| 8      | $\arctan(2^{\wedge}6)$ | 0.015623729 | 0.9 | 10  | 000000001010 | 111111101110 |
| 9      | $\arctan(2^{\wedge}7)$ | 0.007812341 | 0.4 | 5   | 000000000101 | 111111110111 |
| 10     | $\arctan(2^{\wedge}8)$ | 0.003906230 | 0.2 | 3   | 000000000011 | 111111111011 |
| 11     | $\arctan(2^{\wedge}9)$ | 0.001953123 | 0.1 | 1   | 000000000001 | 111111111111 |

with:

RAD = radial  
 DEG = degree  
 DEC = decimal  
 BIN = binary

In the embodiment shown it is checked in the first iteration section IS1 whether the digital 12-bit angle value  $Z_0$  has a positive or a negative sign. To this end the first iteration section IS1 comprises a sign detector SD0 which is coupled to the phase signal input 18 and which detects the sign of the angle value  $Z_0$ , i.e. starting from a two's complement representation it defines the bit value of the most significant bit in the angle value  $Z_0$ . The sign detector SD0 is coupled to control inputs of an inverter circuit I11, an inverter circuit  $\bar{T}12$  which is complementary thereto and an inverter circuit  $\bar{T}13$  which is also complementary. An inverter circuit is hereinafter understood to mean a circuit which multiplies an input signal by the control signal, i.e. it does not invert the input signal when the control signal is +1 and it inverts this signal when the control signal is -1. A complementary inverter circuit works the other way round: signal inversion at a +1 control signal and no signal inversion at a -1 control signal. Signal inputs of I11 and  $\bar{T}12$  are connected to first and second signal inputs 16 and 17, respectively, of the first Cordic 9, while a signal input of  $\bar{T}13$  is coupled to a fixed angle rotation value  $\phi_1$  corresponding to an angle value of  $90^\circ$  as is indicated on the top line of the Table above. A signal output of  $\bar{T}13$  is coupled to a first signal input of an adder A13, while a second signal input of A13 is coupled to the phase signal input 18. Signal outputs of I11,  $\bar{T}12$  and A13 are coupled to delay circuits D12, D11 and D13, respectively.

If the most significant bit in the angle value  $Z_0$  at the phase signal input 18 is positive, the sign of the signal sample  $Y_0$  of the signal input 17 is inverted in the inverter circuit  $\bar{T}12$ , whereafter it is stored as a new signal sample  $Y_1$  in the delay circuit D11, while the signal sample  $X_0$  at the first signal input 16 is stored in the delay circuit D12 as the new signal sample  $X_1$  with the same sign. An opposite sign inversion is effected when the last-mentioned most significant bit in  $Z_0$  is negative. The circuits SD0,  $\bar{T}13$  and A13 constitute the so-called z branch of the iteration section IS1. Not only the sign of the rotation step required for the afore-mentioned processing of the samples  $X_0$  and  $Y_0$  but also a signal processing corresponding to a rotation of the angle value  $Z_1$  through  $90^\circ$  towards the zero value, for example, 0, resulting in  $Z_1 = Z_0 \pm 90^\circ$  is obtained. By representing the angle value in a so-called two's complement form, it will be possible to simplify the circuits required for the angle convergence to a considerable extent and to reduce their size, as will hereinafter be explained in greater detail. The signal processing operations in the first iteration section IS1 described so far are effected in one sampling period, for example,  $ts_1$ . In a subsequent sampling period  $ts_2$  corresponding signal processing operations are performed, starting from the new sample values  $X_1$  and  $Y_1$  and the new angle value  $Z_1$ .

In the second iteration section IS2 the most significant bit value of the angle word stored in D13 is detected in a sign detector SD1 in the last-mentioned sampling period  $ts_2$ . An output of SD1 is connected to control inputs of an inverter circuit I22 and inverter circuits  $\bar{T}21$  and  $\bar{T}23$  which are complementary thereto. Signal inputs of I22 and  $\bar{T}21$  are connected to outputs of D11 and D12, respectively, while a signal input of  $\bar{T}23$  is connected to a fixed angle rotation value  $\phi_2$  corresponding to an angle value of  $45^\circ$ . Signal outputs of  $\bar{T}21$ , I22 and  $\bar{T}23$  are connected to adder circuits A21, A22 and A23, respectively. First signal inputs of A21, A22 and A23 are connected to the signal outputs of D11, D12 and D13, respectively. Signal outputs of A21, A22 and A23 are coupled to delay circuits D21, D22 and D23, respectively.

Dependent on the sign or the value of the most significant bit in the angle value  $Z_1$ , which is supplied from D13 to the z branch (SD1,  $\bar{T}23$  and A23) of the second iteration section S2, the signal of the  $X_1$  and  $Y_1$  signal samples supplied by D11 and D12, respectively, is inverted in the complementary circuit  $\bar{T}21$ , followed by an addition in A21 and A22 with the original value of these signals. The results of these additions are stored as signal samples  $X_2$  and  $Y_2$  in the delay circuits D21 and D22, respectively. In the last-mentioned z branch an angle iteration step is performed again with the aid of the complementary inverter circuit  $\bar{T}23$  and the adder circuit A23, this time through an angle of  $\pi/4$  in

such a direction that the now new angle value Z2 at the output of A23 is smaller than Z1. This new angle value Z2 is subsequently stored in the delay circuit D23.

In a subsequent sampling period  $t_{s3}$  the contents of the delay circuits D21, D22 and D23 are applied as input signal sample for the third iteration section IS3, which has mainly the same circuit configuration as the second iteration section IS2. The third iteration section IS3 comprises a sign detector SD3 an input of which is coupled to an output of the delay circuit D23 and a signal output of which is coupled to control inputs of an inverter circuit I32 and inverter circuits  $\bar{T}31$  and  $\bar{T}33$  which are complementary thereto. Outputs of the delay circuits D21 and D22 are coupled to first signal inputs of adder circuits A31 and A32, respectively, and to signal inputs of I32 and  $\bar{T}31$  via shift registers SH32 and SH31 functioning as divide-by-two circuits. Signal outputs of  $\bar{T}31$  and I32 are coupled to second signal inputs of the adder circuits A31 and A32, respectively. Signal outputs of A31 and A32 are subsequently coupled to signal inputs of delay circuits D31 and D32, respectively. A fixed binary value corresponding to an angle value  $\phi_3$  is applied to a signal input of the complementary inverter circuit  $\bar{T}33$ . Together with the signal output of D23, the signal output of  $\bar{T}33$  is applied to an adder circuit A33 whose signal output is connected to a signal input of a delay circuit D33.

In a corresponding manner an iteration section IS<sub>i</sub> (i = 4 through 11) incorporates inverter circuits I<sub>i2</sub> and inverter circuits  $\bar{T}i1$ ,  $\bar{T}i3$  which are complementary thereto, in combination with adder circuits A<sub>i2</sub>, A<sub>i1</sub> and A<sub>i3</sub> as well as a sign detector SD(i-1) and shift registers SH<sub>i1</sub> and SH<sub>i2</sub>. A division by a factor of  $2^{(i-2)}$  is obtained with the shift registers SH<sub>i1</sub> and SH<sub>i2</sub>. The adder circuits A<sub>i1</sub>, A<sub>i2</sub> and A<sub>i3</sub> are coupled to signal inputs of delay circuits D<sub>i1</sub>, D<sub>i2</sub> and D<sub>i3</sub>, respectively. A signal input of the complementary inverter circuit  $\bar{T}i3$  is connected to a fixed binary value corresponding to an angle value  $\phi_i$ .

The last iteration section IS11 only performs computations with reference to the angle value Z<sub>0</sub> obtained in the penultimate section IS10, which value coincides with zero within a deviation determined by the least significant bit in the angle value word. In the last iteration section IS11 no further angle convergence of Z<sub>10</sub> is performed so that the circuits corresponding to  $\bar{T}i3$  and A<sub>i3</sub>, as well as the delay circuits corresponding to D<sub>i1</sub>, D<sub>i2</sub> and D<sub>i3</sub> are dispensed with.

According to the invention the varying angle value Z<sub>0</sub> is obtained at the phase signal input 18 by means of a sawtooth generator SG. In the version shown this generator comprises an adder circuit A0 with first and second signal inputs and a signal output which is coupled to a signal delay circuit D0, a signal output of which is coupled to the phase signal input 18 of the first Cordic 9 as well as to a second signal input of the adder circuit A0. A variable digital tuning data ft is applied to the first signal input of the adder circuit A0. By a suitable numerical value of the digital tuning data ft with respect to the maximum count of the adder circuit A0, a periodical sawtooth-shaped signal is obtained with angle sampling values Z<sub>0</sub> at the phase signal input 18 in a repetition frequency which corresponds to the RF carrier frequency of the desired RF reception signal.

Fig. 3 shows the successive sampling values of a periodical sawtooth-shaped phase signal z which is obtained in a practical embodiment of a sawtooth generator as is shown in Fig. 2. In this practical embodiment the counting range of the adder circuit A0 is between 0 and 4095 and the maximum count is thus 4095. The numerical value of the digital tuning data is 1031. The Table II below states how the numerical value of the successive samplings or angle values at the phase signal input 18 varies as a function of time.

Tabel II:

| sample |                  |
|--------|------------------|
| 1      | 0                |
| 2      | 1031             |
| 3      | 2062             |
| 4      | 3093             |
| 5      | 28 (4124 - 4096) |
| 6      | 1059             |
| 7      | 2090             |
| 8      | 3121             |
| 9      | 56 (4152 - 4096) |
| 10     | 1087             |
| 11     | 2118             |
| 12     | 3149             |
| 13     | 84 (4180 - 4096) |
| 14     | 1115             |
| 15     | 2146             |
| 16     | 3177             |



Tabel II: (continued)

| sample |                   |
|--------|-------------------|
| 17     | 112 (4208 - 4096) |

Since the maximum count of the adder circuit A0 is a non-integral number of times the numerical values of the digital tuning data, the sampling instants for each sawtooth period shift in phase with respect to the repetition period of the sawtooth. This phase shift results in a frequency spread of the noise so that the noise behaviour of the quadrature mixer stage constituted by the first Cordic 9 and the sawtooth generator SG is considerably improved.

The embodiment shown in Fig. 2 has a so-called pipeline structure because use is made of delay circuits Di1, Di2 and Di3 ( $i = 1 \dots 11$ ). This provides the possibility of performing the consecutive iterations at a clock frequency which is equal to the first sampling frequency. However, it is possible to dispense with the delay circuits Di1, Di2 and Di3. A serial iterative angle convergence in a sampling frequency which is smaller than the clock frequency by a factor equal to the number of iteration sections is then obtained.

Fig. 4 shows a practical implementation of the sawtooth generator SG in combination with the z branch in the cascade circuit of the first to tenth iteration sections IS1 to IS10. Starting from a 12-bit angle value representation, the sawtooth generator SG should comprise twelve parallel accumulator circuits each corresponding to the circuit constituted in Fig. 2 by the adder circuit A0 and the delay circuit D0 and the feedback from the output of the delay circuit D0 to the second signal input of the adder circuit A0. If the accumulator circuit for the least significant bit is referred to as AC1 and that for the least significant bit but one is referred to as AC2, and so forth to AC12 for the accumulator circuit for the most significant bit, the respective circuits AC1 to AC11 should have their carry-out outputs connected to carry-in inputs of the accumulator circuits AC1 to AC12. Signal outputs of the accumulator circuits AC1 to AC11 are connected to delay circuits D131 to D1311, respectively, with a signal inversion being effected in the signal path of the accumulator circuit AC11.

In the case of a two's complement angle value representation as indicated in the Table above, the sign bit S0 for the iteration section IS1 is obtained at the output of the accumulator circuit AC12. The sign bit S1 for the iteration section IS2 is obtained by inverting the output signal of the accumulator circuit AC11 and delaying it over one sampling period in the delay circuit D1311. Also the sign bit S2 for the third iteration section IS3 is solely determined by the output signal of the accumulator circuit AC10.

Signal outputs of the delay circuits D131 to D1310 are connected to signal inputs of delay circuits D231 to D2310, respectively. A signal inversion is effected in the signal path of the accumulator circuit AC10 and the delay circuit D2310 supplies the sign bit S2 to the third iteration section IS3. Outputs of the delay circuits D233 to D239 are connected to signal inputs of first signal inputs of seven adder circuits A233 to A239. The sign bit S2 is applied to second signal inputs of the adder circuits A233, A234, A236 and A239, while the inverted value of the sign bit S2 is applied to the other adder circuits. The adder circuits A233 to A239 are interconnected through a carry-bit coupling, while a carry-bit input of the adder circuit A233 is coupled to the delay circuit D232. The carry-bit output signal of the adder circuit A239 is applied as sign bit S3 to the fourth iteration section IS4 after inversion and delay in a delay circuit D3310. Signal outputs of the adder circuits A233 to A239 are coupled to signal inputs of delay circuits D333 to D339, respectively. Signal outputs of the delay circuits D231 and D232 are coupled to delay circuits D331 and D332, respectively.

For the formation of the fourth sign bit S4 use is made of three adder circuits A337 to A339, first signal inputs of which are coupled to signal outputs of delay circuits D337 to D339, while a second signal input of the adder circuit A338 receives the sign bit S3 and second signal inputs of the adder circuits A337 and A339 receive the inverse value of the sign bit S3. The adder circuits A337 to A339 are interconnected via a carry-bit coupling, while a carry-bit input of the adder circuit A337 is connected to an output of the delay circuit D336. The carry-bit output signal of the adder circuit A339 is applied as sign bit S4 to the fifth iteration section IS5 after inversion and delay in a delay circuit D4310. Signal outputs of the delay circuits D331 to D336 and of the adder circuits A337 to A339 are coupled to delay circuits D431 to D439, respectively.

The sign bit S5 for the sixth iteration section IS6 is obtained by means of eight adder circuits A434 to A439, first inputs of which are coupled to delay circuits D432 to D439, respectively. The input signals of the adder circuits A432 and A436 are inverted. The sign bit S4 is applied to second signal inputs of adder circuits A435 and A437, while the inverse value of the sign bit S4 is applied to the other adder circuits of said group of eight adder circuits A432 to A439. The adder circuits A432 to A439 are interconnected via a carry-bit coupling. The carry-bit input of the adder circuit A432 is connected to an output of the delay circuit D431. Signal outputs of the delay circuit D431 and the adder circuits A432 to A439 are coupled to delay circuits D531 to D539, respectively. An output of the delay circuit D539 supplies the sign bit S5 for the sixth iteration section IS6.

For forming the sign bit S6 for the seventh iteration section IS7 use is made of seven adder circuits A532 to A538, first signal inputs of which are coupled to delay circuits D532 to D538, respectively, while second signal inputs of the

adder circuits A534 and A536 receive the sign bit S5. The inverse value of the sign bit S5 is applied to second signal inputs of the other adder circuits. Also these adder circuits A532 to A538 are interconnected *via* a carry-bit coupling, while a carry-bit input of the adder circuit A532 receives the inverse value of the output signal of the delay circuit D531. Signal outputs of the delay circuit D531 and the adder circuits A532 to A538 are connected to delay circuits D631 to D638, respectively. The output signal of the delay circuit D638 is applied as sign bit S6 to the seventh iteration section IS7.

For forming the sign bit S7 for the eighth iteration section IS8 use is made of four adder circuits A634 to A637, first inputs of which are coupled to outputs of the delay circuits D634 to D637, while a second signal input of the adder circuit A635 receives the sign bit S6. The inverse value of the sign bit S6 is applied to second signal inputs of the other adder circuits A634, A636 and A637. The adder circuits A634 to A637 are interconnected *via* a carry-bit coupling. The delay circuit D642 is coupled to a carry-bit input of the adder circuit A634. Signal outputs of the delay circuits D631 to D633 and adder circuits A634 to A637 are coupled to delay circuits D731 to D737, respectively. The sign bit S7 is obtained at the output of the delay circuit D737.

For forming the sign bit S8 for the ninth iteration section IS9 use is made of four adder circuits A733 to A736. First signal inputs of the adder circuits A733 to A736 are coupled to delay circuits D733 to D737, respectively. A signal inversion is effected at the first signal input of the adder circuit A733. A second signal input of the adder circuit A734 receives the sign bit S7, while second signal inputs of the adder circuits A733, A735 and A736 receive the inverse value of the sign bit S7. The adder circuits A733 to A736 are interconnected *via* a carry-bit coupling, while a carry-bit input of the adder circuit A733 is coupled to an output of the delay circuit D732. The delay circuits D731, D732 and the adder circuits A733 to A736 are connected to delay circuits D831 to D836, respectively. The delay circuit D836 supplies said sign bit S8.

For forming the sign bit S9 for the tenth iteration section IS10 use is made of four adder circuits A832 to A835, signal inputs of which are coupled to outputs of the delay circuits D832 to D835. A signal inversion is effected at the first signal input of the adder circuit A832. A second signal input of the adder circuit A833 receives the sign bit S8, while second signal inputs of the adder circuits A832, A834 and A835 receive the inverse value of the sign bit S8. The sign bit S9 for the iteration section S10 is subsequently derived from the output signal of the adder circuit A835 *via* a delay circuit D935.

The circuit shown in Fig. 4 and described hereinbefore comprises a minimum number of circuit components with which the functions of the sawtooth generator SG and the z branch in Fig. 2 required for generating the sign bits S0 to S10 are realised. This is possible by choosing the  $\phi$  values in the iteration sections IS1 to IS10 in accordance with the Table I above. As already stated hereinbefore, the angle value  $Z_0$  supplied by the sawtooth generator SG is iteratively rotated to a residual or zero value. For each iteration section the variation range of  $Z_i$  decreases. Thus, it is not necessary to pass on all bit values at the side of the most significant bit from the output to a subsequent iteration section. As a result the word width of the angle value  $z$  in which variations may occur decreases for each iteration section. This yields a first economy of circuit components. Optimum use is further made of so-called wiring logic techniques, or realising logic functions by means of suitably chosen through-connections.

Fig. 5 shows a block diagram of a practical embodiment of the RF digital filter 6 in which the digital RF signal at the output of the A/D converter 5 is split in phase into a pair of phase quadrature RF signals  $x_i$  and  $y_i$ . The RF digital filter shown is a so-called anti-symmetrical FIR (Finite Impulse Response) filter and is provided with a series circuit of six delay circuits D1 to D6, the input of D1 and the outputs of D2, D4 and D6 being coupled to multiplier circuits M1 to M4, respectively. Outputs of these multiplier circuits are connected to inputs of an adder circuit AD. The input sampling value of the cascade circuit and the output sampling value of the delay circuits D2, D4 and D6 are multiplied by means of the multiplier circuits M1 to M4 by fixed weighting factors which are pairwise mirrored with respect to 0 and are -0.0625; -0.5625; 0.5625; 0.0625, respectively. Starting from a digital RF input signal at the common input of M1 and D1 the digital filter shown applies a digital RF signal  $x_i$  to the output of D3, which signal is applied to the first signal input 16 of the first Cordic 9 and is in an accurate phase quadrature relationship with the output signal of the adder AD which is applied as RF digital signal  $y_i$  to the second signal input of the first Cordic 9. The signal delay which occurs between the input of the filter and the output of D3 is equal to the signal delay between said input and the output of the adder circuit AD.

It is possible to use an anti-symmetrical digital filter of an order which is higher than that of the one shown, for which it generally holds that such a filter is provided with a series circuit of  $2n+1$  ( $n = 0, 1 \dots$ ) serial delay circuits, the output of each pair of delay circuits as well as the input of the series circuit being coupled to an adder circuit *via* a weighting factor multiplier circuit, while the common connection between the delay circuit of the  $(n+1)^{\text{th}}$  pair is coupled to the first signal input and the output of the adder circuit is coupled to the second signal input of the first Cordic 9.

However, it is very well possible to perform the combined functions of signal delay compensation and Hilbert transformation separately by means of the digital filter shown, *i.e.* by means of mutually separated circuits. As stated hereinbefore, such a circuit can be dispensed with by adapting the selectivity after the first Cordic and the output signal of the A/D converter can be directly applied to either the first signal input 16 or the second signal input 17 of the first

Cordic 9.

Fig. 6 shows a sub-sample or decimating low-pass filter section for use in each of the filters 10 and 11 of the digital filter device 10, 11 of the receiver of Fig. 1.

The filter section shown comprises a cascade circuit of a first-order comb filter D1, AD1, a second-order comb filter D2, AD2, an accumulation and dump circuit D3, AD3 provided with adder circuits AD1, AD2 and AD3, respectively, each with first and second signal inputs and a signal output. The first signal input of AD1 constitutes a signal input IN of the filter section and is coupled to the second signal input of AD1 via a delay circuit D1, while the signal output of AD1 is connected to the first signal input of AD2 and to the second signal input of AD2 via a delay circuit D2, the signal output of AD2 is connected to a first signal input of AD3 and the signal output of AD3 is connected to the second signal input of AD3 via a delay circuit D3 and also constitutes a signal output OUT of the filter section. The comb filter D1, AD1 and D2, AD2 is of the first and second order, respectively, because D1 and D2 realise a signal delay over one period and two periods, respectively, of the sampling frequency fsi of the input signal applied to IN. Likewise as D1, the delay circuit D3 realises a signal delay over one period of fsi and is preferably realised by means of a resettable memory circuit.

The output signal of the accumulation and dump circuit AD3, D3 is sampled at an output sampling frequency fso in which, for example,  $f_{so} = 1/4 f_{si}$ , while directly after each output signal sampling D3 the entire accumulation and dump circuit AD3, D3 is thus reset. As a result a sampling frequency reduction or a decimation by a factor of 4 is effected. Simultaneously a selectivity is realised which is comparable with that of a time-invariant filter having a transfer of  $H(z) = (1+z^{-1})(1+z^{-2})$ . No sampling frequency decimation is effected in such a time-invariant filter, while the realisation of such a filter requires considerably more components than the accumulation and dump circuit D3, AD3.

To prevent noise and other unwanted interference components from aliasing as a result of the decimation in the accumulation and dump circuit AD3, D3 in the frequency band of the useful signal, these unwanted interference components are firstly suppressed by means of the comb filters AD1, D1 and AD2, D2 in a sufficiently wide cut-off range around the zeros in the filter transfer. The comb filters AD1, D1 and AD2, D2 jointly have a transfer which is equal to  $(1+z^{-1})$  and  $(1+z^{-2})$ .

In a practical embodiment of the receiver of Fig. 1 as a directly mixing FM receiver suitable for receiving RF-FM signals in a range between 87.5 and 108 MHz, a first sampling frequency fsi of 350 MHz was chosen in connection with avoiding a 3rd harmonic interference in the desired signal at the A/D conversion in the A/D converter 5 and obtaining a simple realisation of the digital RF quadrature filter 6.

For each of the two phase quadrature baseband signals xo and yo at the outputs 19 and 20 of the first Cordic 9 a cascade circuit of three filter sections of the type shown in Fig. 6 and described hereinbefore was used, while a first-order and a second-order comb filter corresponding to D1, AD1 and D2, AD2 (not shown) were added as extra components to the third and last filter section in the cascade circuit. An extra suppression of the unwanted interference components and a comparatively small decrease of the number of circuit elements was obtained with these extra comb filters. A decimation factor of  $4^3 = 64$  was obtained with the last-mentioned cascade circuit of filter sections, so that starting from  $f_{si} = 350$  MHz an output sampling frequency of approximately 5.47 MHz was obtained. With this last-mentioned filter section cascade circuit a selectivity was realised which is comparable with that of a time-invariant comb filter having a transfer function of  $(1+z^{-1})^2(1+z^{-2})^2(1+z^{-4})^2(1+z^{-8})^2(1+z^{-16})^3(1+z^{-32})^3$ , and such a time-invariant comb filter (not shown) should then be succeeded by a decimation device and requires a considerably larger number of circuit components than said filter section cascade circuit.

In the relevant practical embodiment each of the two filter section cascade circuits was succeeded by a selective low-pass filter of the aforementioned "wave digital filter" type at whose output a last sampling frequency decimation by a factor of 5 or 6 can be effected.

The order and the number of comb filters in each filter section is to be chosen in dependence upon the required receiver specifications, the spectral width and location of the desired signal to be filtered and/or the decimation factor. For example, it is very well possible to dispense with one or even with said two comb filters in one or more filter sections, or it may be necessary to add extra comb filters.

## Claims

1. A receiver comprising an A/D converter (5) for digitally sampling an analog signal modulated on a carrier frequency at a first sampling frequency, consecutively coupled to a digital quadrature mixer stage (6,9) for a carrier frequency shift of the digitized modulated signal from the A/D converter (5), the digital quadrature mixer stage being provided with first and second signal outputs (19,20) via which it supplies a pair of carrier frequency converted phase quadrature signals, a digital filter device (10,11) for selecting the phase quadrature signals of the quadrature mixer stage and for decimating the sampling frequency from said first sampling frequency to a second sampling frequency.

cy, and a digital demodulation device (12,13) being coupled to the digital filter device, characterized in that the digital quadrature mixer stage comprises a first coordinate rotation digital computer (Cordic) (9) in the rotation mode, having at least a first signal input (16,17) which is coupled to an output of the A/D converter (5) and a phase signal input (18) which is coupled to an output of a digital sawtooth generator (SG), which digital sawtooth generator supplies a periodical digital sawtooth-shaped phase signal to the phase signal input of said computer for a periodical  $2\pi$  phase rotation of at least the signal applied to the computer *via* the first signal input at a repetition frequency which is equal to the magnitude of said carrierfrequency shift, said computer supplying at the first and second signal outputs (19,20) said pair of carrierfrequency converted phase quadrature signals to the digital filter device (10,11).

2. A receiver as claimed in Claim 1, characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal is variable in a range which corresponds to the frequency reception range of the receiver.
3. A receiver as claimed in Claim 1 or 2, characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal of the digital sawtooth generator is equal to said carrier frequency of the analog signal applied to the A/D converter for a direct frequency conversion in the computer to the baseband.
4. A receiver as claimed in Claim 1 or 2, characterized in that the repetition frequency of the periodical digital sawtooth-shaped phase signal deviates by a fixed intermediate frequency value from said carrier frequency of the analog signal applied to the A/D converter.
5. A receiver as claimed in any one of the preceding Claims, characterized in that the phase rotation caused by the periodical digital sawtooth-shaped phase signal varies linearly over a range of  $2\pi$  radials in each period.
6. A receiver as claimed in any one of the preceding Claims, characterized in that the first sampling frequency is a non-integral number of times the carrierfrequency of the analog signal applied to the A/D converter, which number is larger than two.
7. A receiver as claimed in Claim 6, characterized in that the digital sawtooth generator comprises an accumulator circuit which is provided with a cascade circuit of an adder circuit having first and second signal inputs and a delay circuit for a signal delay over one sampling period, an output of the delay circuit being coupled to the phase signal input of the first computer as well as to the second signal input of the adder circuit, a digital tuning data being applied to said first signal input of the adder circuit, the magnitude of said tuning data being variable and being included in the counting range of the adder circuit by a non-integral number of times which is larger than two.
8. A receiver as claimed in any one of the preceding Claims, characterized in that the digital demodulation device comprises a second coordinate rotation digital computer (CORDIC) in the vectoring mode having first and second signal inputs to which the selected phase quadrature signals of the digital filter device at the second sampling frequency are applied, and an output which is coupled to a signal reproducing device.
9. A receiver as claimed in Claim 8, for receiving FM signals, characterized in that said output of the second coordinate rotation digital computer (CORDIC) is the phase signal output and is coupled to the signal reproducing device *via* a differentiating circuit.
10. A receiver for receiving FM signals as claimed in Claim 9, characterized by an RF input section coupled between an antenna input and an input of the A/D converter, comprising at least an RF filter device for a bandpass selection mainly between 85 MHz and 110 MHz, an automatically controlled amplifier for adapting the signal dynamic range of the reception signal to the dynamic range of the A/D converter, the first sampling frequency of the A/D converter being of the order of 350 MHz.
11. A receiver as claimed in Claim 8, for receiving AM signals, characterized in that said output is a first signal output of the second coordinate rotation digital computer (Cordic) to which an output signal is supplied which varies with the amplitude of the input signals of said second computer, said first signal output being coupled to the signal reproducing device *via* an amplitude detector.
12. A receiver as claimed in any one of the preceding Claims, characterized in that the first coordinate rotation digital computer has a second signal input, while a Hilbert signal transformation is effected between the A/D converter and one of the two signal inputs and a signal delay compensation is effected between the A/D converter and the

other signal input.

13. A receiver as claimed in Claim 12, characterized in that a digital anti-symmetrical finite impulse response filter is incorporated between the A/D converter on the one hand and the first and second signal inputs of the first coordinate rotation digital computer on the other hand, which filter comprises a series arrangement of  $2n+1$  ( $n = 0, 1, \dots$ ) serial pairs of delay circuits, the output of each pair of delay circuits as well as the input of the series circuit being coupled to an adder circuit via a weighting factor multiplier circuit, the common connections between the delay circuits of the  $(n+1)^{\text{th}}$  pair being coupled to the first signal input and the output of the adder circuit being coupled to the second signal input of the first coordinate rotation digital computer.

14. A receiver as claimed in Claim 7, characterized in that the first coordinate rotation digital computer (Cordic) comprises a cascade circuit of first to  $k^{\text{th}}$  iteration sections each comprising first and second phase quadrature signal branches and a phase signal branch coupled between the first signal input and output, the second signal input and output and the phase signal input and output of said first computer, said phase signal branch in each of the first to  $(k-1)^{\text{th}}$  iteration sections comprising a sign detector for detecting the sign of the most significant information bit in the angle value applied to the phase signal branch of the relevant iteration section, as well as an adder circuit to which a fixed rotation angle added to or subtracted from the angle value is applied in dependence upon the last-mentioned sign, said adder circuit of the first to  $(k-1)^{\text{th}}$  iteration section being coupled to the phase signal branch of the second to  $k^{\text{th}}$  iteration section, the fixed rotation angle in the first and second iteration sections being  $90^\circ$  and  $45^\circ$ , respectively, and the tangential value of the fixed rotation angles in the 3rd to  $(k-1)^{\text{th}}$  iteration sections being equal to  $2^{-i}$  with  $i = 1$  to  $k-3$ .

15. A receiver as claimed in Claim 14, characterized in that the angle value and the fixed rotation angle are each represented in a two's complement mode word and in that a binary adder circuit is incorporated in each iteration section only for those bit sites in the angle value word at which the bit value can vary as a result of the addition or subtraction to or from the fixed rotation angle.

16. A receiver as claimed in any one of the preceding Claims, characterized in that said digital filter device has a sub-sample filter section which comprises a cascade circuit of an  $n^{\text{th}}$  order comb filter and an accumulation and dump circuit comprising an adder circuit having first and second signal inputs and a signal output, said first signal input being coupled to an output of the comb filter device and said signal output being coupled to the second signal input via a delay circuit for a signal delay of an input sampling frequency over one period, the output signal of the adder circuit being sampled at an output sampling frequency which is  $1/(2n)$  the input sampling frequency and the accumulation and dump circuit being reset after each last-mentioned sampling.

17. A receiver as claimed in Claim 16, characterized in that the  $n^{\text{th}}$  order comb filter is arranged in cascade with  $n-1$  further comb filters of the order 1 to  $n-1$ , respectively.

#### Patentansprüche

1. Empfänger mit einem Analog-Digital-Wandler (5) für eine digitale Abtastung eines einer Trägerfrequenz aufmodulierten analogen Signals mit einer ersten Abtastfrequenz, wobei dieser Wandler nacheinander gekoppelt ist mit einer digitalen Quadraturmischstufe (6, 9) für eine Trägerfrequenzverschiebung des digitalisierten modulierten Signals des Analog-Digital-Wandlers (5), mit einem ersten und einem zweiten Signalausgang (19, 20), über welche die digitale Quadraturmischstufe ein Paar in Trägerfrequenz umgewandelter Phasenquadratursignale liefert, mit einer digitalen Filteranordnung (10, 11) für eine Selektion der Phasenquadratursignale der Quadraturmischstufe sowie für eine Dezimierung der Abtastfrequenz von der genannten ersten zu einer zweiten Abtastfrequenz und mit einer digitalen Demodulationsanordnung (12, 13), die mit der digitalen Filteranordnung gekoppelt ist, dadurch gekennzeichnet, daß die digitale Quadraturmischstufe eine erste digitale Koordinaten-Rotations-Rechenanordnung (Cordic) (9) in dem Rotationsmodus aufweist, die mit wenigstens einem ersten Signaleingang (16, 17) versehen ist, der mit einem Ausgang des Analog-Digital-Wandlers (5) gekoppelt ist, und mit einem Phasensignaleingang (18) versehen ist, der mit einem Ausgang eines digitalen Sägezahngenerators (SG) gekoppelt ist, der dem Phasensignaleingang der genannten Rechenanordnung ein periodisches digitales sägezahnförmiges Phasensignal liefert für eine periodische  $2\pi$ -Phasendrehung wenigstens des über den ersten Signaleingang der Rechenanordnung zugeführten Signals mit einer Wiederholungsfrequenz, die der Größe der genannten Trägerfrequenzverschiebung entspricht, wobei die genannte Rechenanordnung an ihren ersten und zweiten Signalausgängen (19, 20) der digitalen Filteranordnung (10, 11) das genannte Paar in der Trägerfrequenz umgewandelter Phasenqua-

dratursignale liefert.

2. Empfänger nach Anspruch 1, dadurch gekennzeichnet, daß die Wiederholungsfrequenz des periodischen digitalen sägezahnförmigen Phasensignals in einem Bereich veränderlich ist, der dem Frequenzempfangsbereich des Empfängers entspricht.
3. Empfänger nach Anspruch 1 oder 2, dadurch gekennzeichnet, daß die Wiederholungsfrequenz des periodischen digitalen sägezahnförmigen Phasensignals des digitalen Sägezahngenerators der genannten Trägerfrequenz des analogen Signals entspricht, das dem Analog-Digital-Wandler für eine direkte Frequenzumwandlung in der Rechenanordnung zu dem Basisband zugeführt wird.
4. Empfänger nach Anspruch 1 oder 2, dadurch gekennzeichnet, daß die Wiederholungsfrequenz des periodischen digitalen sägezahnförmigen Phasensignals um einen festen ZF-Wert von der genannten Trägerfrequenz des dem Analog-Digital-Wandler zugeführten analogen Signals abweicht.
5. Empfänger nach einem der vorstehenden Ansprüche, dadurch gekennzeichnet, daß die durch das periodische digitale sägezahnförmige Phasensignal verursachte Phasendrehung in jeder Periode über einen Bereich von  $2\pi$  rad linear variiert.
6. Empfänger nach einem der vorstehenden Ansprüche, dadurch gekennzeichnet, daß die erste Abtastfrequenz eine nicht ganze Anzahl Male größer ist als die Trägerfrequenz des dem Analog-Digital-Wandler zugeführten analogen Signals, wobei die Anzahl größer ist als zwei.
7. Empfänger nach Anspruch 6, dadurch gekennzeichnet, daß der digitale Sägezahngenerator eine Speicherschaltung aufweist, die mit einer Kaskadenschaltung einer Addierschaltung mit einem ersten und einem zweiten Signaleingang und mit einer Verzögerungsschaltung versehen ist für eine Signalverzögerung über eine Abtastperiode, wobei ein Ausgangssignal der Verzögerungsschaltung mit dem Phasensignaleingang der ersten Rechenanordnung sowie mit dem zweiten Signaleingang der Addierschaltung gekoppelt ist, wobei dem genannten ersten Signaleingang der Addierschaltung eine digitale Abstimminformation zugeführt wird, wobei die Größe der genannten Abstimminformation veränderlich ist und durch eine nicht ganze Anzahl Male in dem Zählbereich der Addierschaltung eingeschlossen, wobei diese Anzahl größer ist als zwei.
8. Empfänger nach einem der vorstehenden Ansprüche, dadurch gekennzeichnet, daß der digitale Demodulator mit einer zweiten Cordic in dem Vectoring-Modus versehen ist mit ersten und zweiten Signaleingängen, denen die selektierten Phasenquadratursignale der digitalen Filteranordnung in der zweiten Abtastfrequenz zugeführt worden sind und mit einem Ausgang, der mit einer Signalwiedergabeordnung gekoppelt ist.
9. Empfänger nach Anspruch 8 zum Empfangen von FM-Signalen, dadurch gekennzeichnet, daß der genannte Ausgang der CORDIC der Phasensignalausgang ist und über eine Differenzierschaltung mit der Signalwiedergabeordnung gekoppelt ist.
10. Empfänger für den Empfang von FM-Signalen nach Anspruch 9, gekennzeichnet durch einen zwischen einem Antenneingang und einem Eingang des Analog-Digital-Wandlers vorgesehenen HF-Eingangsteil mit wenigstens einer HF-Filteranordnung für eine Bandpaßselektion von im wesentlichen 85 MHz bis 110 MHz, einem automatisch geregelten Verstärker für eine Anpassung der Signaldynamik des Empfangssignals an den dynamischen Bereich des Analog-Digital-Wandlers, wobei die erste Abtastfrequenz des Analog-Digital-Wandlers von der Größenordnung von 350 MHz ist.
11. Empfänger nach Anspruch 8 für den Empfang von AM-Signalen, dadurch gekennzeichnet, daß der genannte Ausgang einen ersten Signalausgang der zweiten CORDIC ist, an dem ein Ausgangssignal geliefert wird, das mit der Amplitude der Eingangssignale dieser zweiten Rechenanordnung schwankt, wobei dieser erste Signalausgang über einen Amplitudendetektor mit der Signalwiedergabeordnung gekoppelt ist.
12. Empfänger nach einem der vorstehenden Ansprüche, dadurch gekennzeichnet, daß die erste CORDIC mit einem zweiten Signaleingang versehen ist, wobei zwischen dem Analog-Digital-Wandler und einem der beiden Signaleingänge eine Hilbert-Signaltransformation stattfindet und zwischen dem Analog-Digital-Wandler und dem anderen Signaleingang ein Signalverzögerungsausgleich erfolgt.

13. Empfänger nach Anspruch 12, dadurch gekennzeichnet, daß zwischen dem Analog-Digital-Wandler einerseits und den ersten und zweiten Signaleingängen der ersten Cordic andererseits ein digitales antisymmetrisches FIR-Filter vorgesehen ist, das mit einer Reihenschaltung aus  $2n+1$  ( $n = 0, 1 \dots$ ) seriellen Paaren von Verzögerungsschaltungen versehen ist, wobei der Ausgang jedes Paares von Verzögerungsschaltungen sowie der Eingang der Reihenschaltung über eine Gewichtungsfaktor-Multiplizierschaltung mit einer Addierschaltung gekoppelt sind, wobei die gemeinsamen Verbindungen zwischen der Verzögerungsschaltung und dem  $(n+1)$ . Paar mit dem ersten Signaleingang gekoppelt ist und der Ausgang der Addierschaltung mit dem zweiten Signaleingang der ersten Cordic gekoppelt ist.
14. Empfänger nach Anspruch 7, dadurch gekennzeichnet, daß die erste CORDIC eine Kaskadenschaltung erster bis einschließlich  $k$ . Iterationsteile aufweist, die je mit einem ersten und einem zweiten Phasenquadratursignalzweig und einem Phasensignalzweig versehen sind, die zwischen dem ersten Signaleingang und -ausgang, dem zweiten Signaleingang und -ausgang bzw. dem Phasensignaleingang und -ausgang der genannten ersten Rechenanordnung vorgesehen sind, wobei der Phasensignalzweig in jedem der ersten bis einschließlich  $(k-1)$ . Iterationsteile einen Vorzeichendetektor aufweist zur Detektion des Vorzeichens des signifikantesten Informationsbits in dem dem Phasensignalzweig des betreffenden Iterationsteils zugeführten Winkelwert, sowie eine Addierschaltung, der in Abhängigkeit von dem letztgenannten Vorzeichen ein fester Rotationswinkel zu dem Winkelwert hinzuaddiert bzw. von demselben subtrahiert wird, wobei diese Addierschaltung des ersten bis einschließlich  $(k-1)$ . Iterationsteils mit dem Phasensignalzweig des zweiten bis einschließlich  $k$ . Iterationsteil gekoppelt ist, wobei der feste Rotationswinkel in dem ersten und zweiten Iterationsteil  $90^\circ$  bzw.  $45^\circ$  beträgt und der Tangenswert der festen Rotationswinkel in dem 3. bis einschließlich  $(k-1)$ . Iterationsteil dem Wert  $2^{-i}$  beträgt, mit  $i = 1$  bis einschließlich  $k-3$ .
15. Empfänger nach Anspruch 14, dadurch gekennzeichnet, daß der Winkelwert des festen Rotationswinkels je in einem "two's complement"-Moduswort dargestellt ist und in jedem Iterationsteil eine binäre Addierschaltung vorgesehen ist, nur für diejenigen Bitstellen in dem Winkelwort, an denen der Bitwert infolge der Addition oder Subtraktion von dem festen Rotationswinkel abweichen kann.
16. Empfänger nach einem der vorstehenden Ansprüche, dadurch gekennzeichnet, daß die genannte digitale Filteranordnung mit einem Unterabstastfilterteil versehen ist, der eine Kaskadenschaltung eines Kammfilters der  $n$ . Ordnung und einer Akkumulierung und Schutzschaltung aufweist, die mit einer Addierschaltung mit ersten und zweiten Signaleingängen und mit einem Signalausgang versehen ist, wobei der erste Signaleingang mit einem Ausgang der Kammfilteranordnung gekoppelt ist und wobei der Signalausgang über eine Verzögerungsschaltung für eine Signalverzögerung um nur eine Periode einer Eingangsabstastfrequenz mit diesem zweiten Signaleingang gekoppelt ist, wobei eine Abtastung des Ausgangssignals der Addierschaltung mit einer Ausgangsabstastfrequenz, die  $1/(2n)$  mal der Eingangsabstastfrequenz beträgt, und die Akkumulierung- und Schutzschaltung nach jeder letztgenannten Abtastung rückgestellt wird.
17. Empfänger nach Anspruch 16, dadurch gekennzeichnet, daß das Kammfilter  $n$ . Ordnung mit  $n-1$  weiteren Kammfiltern der 1. bis einschließlich  $(n-1)$ . Ordnung in Kaskade geschaltet ist.

#### Revendications

1. Récepteur comprenant un convertisseur A/D (5) pour échantillonner numériquement un signal analogique modulé sur une fréquence porteuse à une première fréquence d'échantillonnage, couplé ensuite à un étage mélangeur numérique en quadrature (6, 9) pour assurer un décalage de fréquence porteuse du signal numérisé modulé provenant du convertisseur A/D (5), ledit étage mélangeur numérique en quadrature étant pourvu d'une première et d'une deuxième sorties de signaux (19, 20) par lesquelles il délivre deux signaux en quadrature de phase ayant subi une conversion de fréquence porteuse, un dispositif de filtrage numérique (10, 11) pour sélectionner les signaux en quadrature de phase de l'étage mélangeur en quadrature et pour décimer la fréquence d'échantillonnage depuis ladite première fréquence d'échantillonnage jusqu'à une deuxième fréquence d'échantillonnage, et un dispositif de démodulation numérique (12, 13) étant couplé au dispositif de filtrage numérique, caractérisé en ce que l'étage mélangeur en quadrature numérique comprend un premier ordinateur numérique à rotation de coordonnées (Cordic) (9) dans le mode de rotation, ayant au moins une première entrée de signaux (16, 17) qui est couplée à une sortie du convertisseur A/D (5) et une entrée de signaux de phase (18) qui est couplée à une sortie d'un générateur de signaux en dents de scie numériques (SG), ledit générateur de signaux en dents de scie numériques délivrant un signal de phase en dents de scie numérique périodique à l'entrée de signaux de phase dudit ordinateur pour une rotation de phase périodique de  $2\pi$  d'au moins le signal appliqué à l'ordinateur via la

première entrée de signaux à une fréquence de répétition qui est égale à la grandeur dudit décalage de la fréquence porteuse, ledit ordinateur délivrant, à la première et à la deuxième sorties de signaux (19, 20), lesdits deux signaux en quadrature de phase ayant subi une conversion de fréquence porteuse au dispositif de filtrage numérique (10, 11).

2. Récepteur selon la revendication 1, caractérisé en ce que la fréquence de répétition du signal de phase en dents de scie numérique périodique peut varier dans une plage qui correspond à la plage de réception de fréquence du récepteur.
3. Récepteur selon la revendication 1 ou 2, caractérisé en ce que la fréquence de répétition du signal de phase en dents de scie numérique périodique du générateur de signaux en dents de scie numériques est égale à ladite fréquence porteuse du signal analogique appliqué au convertisseur A/D pour assurer une conversion de fréquence directe dans l'ordinateur à la bande de base.
4. Récepteur selon la revendication 1 ou 2, caractérisé en ce que la fréquence de répétition du signal de phase en dents de scie numérique périodique dévie d'une valeur de fréquence intermédiaire fixe de ladite fréquence porteuse du signal analogique appliqué au convertisseur A/D.
5. Récepteur selon l'une quelconque des revendications précédentes, caractérisé en ce que la rotation de phase provoquée par le signal de phase en dents de scie numérique périodique varie linéairement sur une plage de  $2\pi$  radians dans chaque période.
6. Récepteur selon l'une quelconque des revendications précédentes, caractérisé en ce que la première fréquence d'échantillonnage vaut un nombre de fois non entier la fréquence porteuse du signal analogique appliqué au convertisseur A/D, ledit nombre étant supérieur à deux.
7. Récepteur selon la revendication 6, caractérisé en ce que le générateur de signaux en dents de scie comprend un circuit accumulateur qui est pourvu d'un montage en cascade d'un circuit additionneur ayant une première et une deuxième entrées de signaux et d'un circuit à retard pour un retardement du signal sur une période d'échantillonnage, la sortie du circuit à retard étant couplée à l'entrée de signaux de phase du premier ordinateur ainsi qu'à la deuxième entrée de signaux du circuit additionneur, les données d'accord numériques étant appliquées à ladite première entrée de signaux du circuit additionneur, la grandeur desdites données d'accord étant variable et étant incluse dans la plage de comptage du circuit additionneur par un nombre non entier de fois qui est supérieur à deux.
8. Récepteur selon l'une quelconque des revendications précédentes, caractérisé en ce que le dispositif de démodulation numérique comprend un deuxième ordinateur numérique à rotation de coordonnées (Cordic) en mode vectoriel, ayant une première et une deuxième entrées de signaux auxquelles les signaux en quadrature de phase sélectionnés du dispositif de filtrage numérique à la deuxième fréquence d'échantillonnage sont appliqués, et une sortie qui est couplée à un dispositif de reproduction de signaux.
9. Récepteur selon la revendication 8, pour recevoir des signaux FM, caractérisé en ce que ladite sortie du deuxième ordinateur numérique à rotation de coordonnées (Cordic) est la sortie de signaux de phase et est couplée au dispositif de reproduction de signaux via un circuit de différenciation.
10. Récepteur pour recevoir des signaux FM selon la revendication 9, caractérisé par une section d'entrée RF couplée entre une entrée d'antenne et une sortie du convertisseur A/D, comprenant au moins un dispositif de filtrage RF pour une sélection de bande passante principalement entre 85 MHz et 110 MHz, un amplificateur commandé automatiquement pour adapter la plage dynamique du signal de réception à la plage dynamique du convertisseur A/D, la première fréquence d'échantillonnage du convertisseur A/D étant de l'ordre de 350 MHz.
11. Récepteur selon la revendication 8 pour recevoir des signaux AM, caractérisé en ce que ladite sortie est une première sortie de signaux du deuxième ordinateur numérique à rotation de coordonnées (Cordic) auquel est délivré un signal de sortie qui varie avec l'amplitude des signaux d'entrée dudit deuxième ordinateur, ladite première sortie de signaux étant couplée au dispositif de reproduction de signaux via un détecteur d'amplitude.
12. Récepteur selon l'une quelconque des revendications précédentes, caractérisé en ce que le premier ordinateur numérique à rotation de coordonnées a une deuxième entrée de signaux, tandis qu'une transformation de signaux



de Hilbert est effectuée entre le convertisseur A/D et l'une des deux entrées de signaux et qu'une compensation du retard de signaux est effectuée entre le convertisseur A/D et l'autre entrée de signaux.

- 5 13. Récepteur selon la revendication 12, caractérisé en ce qu'un filtre à réponse impulsionnelle finie antisymétrique numérique est incorporé entre le convertisseur A/D, d'une part, et la première et la deuxième entrées de signaux du premier ordinateur numérique à rotation de coordonnées, d'autre part, ledit filtre comprenant un montage en série de  $2n+1$  ( $n = 0, 1, \dots$ ) paires sérielles de circuits à retard, la sortie de chaque paire de circuits à retard ainsi que l'entrée du montage en série étant couplées à un circuit additionneur via un circuit multiplicateur par un facteur de pondération, les connexions communes entre le circuit à retard de la  $(n + 1)^{\text{ème}}$  paire étant couplées à la première entrée de signaux et la sortie du circuit additionneur étant couplée à la deuxième entrée de signaux du premier ordinateur numérique à rotation de coordonnées.
- 10 14. Récepteur selon la revendication 7, caractérisé en ce que le premier ordinateur numérique à rotation de coordonnées (Cordic) comprend un montage en cascade d'une première à une  $k^{\text{ème}}$  sections d'itération comprenant chacune une première et une deuxième branches de signaux en quadrature de phase et une branche de signaux de phase couplées entre la première entrée de signaux et la sortie, la deuxième entrée de signaux et l'entrée et la sortie de signaux de phase dudit premier ordinateur, ladite branche de signaux de phase de chacune de la première à la  $(k - 1)^{\text{ème}}$  sections d'itération comprenant un détecteur de signe pour détecter le signe du bit d'information le plus significatif dans la valeur d'angle appliquée à la branche de signaux de phase de la section d'itération concernée, ainsi qu'un circuit additionneur auquel un angle de rotation fixe ajouté ou soustrait de la valeur d'angle est appliqué en fonction du dernier signe mentionné, ledit circuit additionneur de la première à la  $(k - 1)^{\text{ème}}$  sections d'itération étant couplé à la branche de signaux de phase de la deuxième à la  $k^{\text{ème}}$  sections d'itération, l'angle de rotation fixe de la première et de la deuxième sections d'itération étant de  $90^\circ$  et  $45^\circ$ , respectivement, et la valeur tangentielle des angles de rotation fixes de la troisième à la  $(k - 1)^{\text{ème}}$  sections d'itération étant égale à  $2^{-i}$  avec  $i = 1$  à  $k - 3$ .
- 15 20 25 15. Récepteur selon la revendication 14, caractérisé en ce que la valeur d'angle et l'angle de rotation fixe sont chacun représentés dans un mot de mode avec complément à deux et en ce qu'un circuit additionneur binaire est incorporé à chaque section d'itération uniquement pour les sites binaires du mot de valeur d'angle pour lesquels la valeur binaire peut varier suite à l'addition ou à la soustraction de l'angle de rotation fixe.
- 30 35 40 16. Récepteur selon l'une quelconque des revendications précédentes, caractérisé en ce que ledit dispositif de filtrage numérique a une section de filtre de sous-échantillonnage qui comprend un montage en cascade d'un filtre en peigne du  $n^{\text{ème}}$  ordre et un circuit d'accumulation et de vidage comprenant un circuit additionneur ayant une première et une deuxième entrées de signaux et une sortie de signaux, ladite première entrée de signaux étant couplée à une sortie du dispositif de filtrage en peigne et ladite sortie de signaux étant couplée à la deuxième entrée de signaux via un circuit à retard pour retarder les signaux d'une fréquence d'échantillonnage d'entrée d'une période, le signal de sortie du circuit additionneur étant échantillonné à une fréquence d'échantillonnage de sortie qui est de  $1/(2n)$  de la fréquence d'échantillonnage d'entrée et le circuit d'accumulation et de vidage étant remis à zéro après chacun de ces échantillonnages mentionnés.
- 45 50 55 17. Récepteur selon la revendication 16, caractérisé en ce que le filtre en peigne du  $n^{\text{ème}}$  ordre est agencé en cascade avec  $n-1$  autres filtres en peigne d'ordres 1 à  $n-1$ , respectivement.

The graph displays the variation of the zero-order Bessel function  $Z_0$  (in LIN) against the sample number (in LIN). The vertical axis ( $Z_0$ ) ranges from 0.0 to 4.5E3, with major ticks every 500.0 units. The horizontal axis (Sample number) ranges from 0 to 16, with major ticks every 2 units. The plot shows a periodic sawtooth pattern with sharp vertical spikes reaching a maximum value of approximately 4.0E3. The dashed line represents the envelope of the function, which is a periodic sawtooth wave. The solid line represents the actual function, which follows the envelope but has sharp vertical spikes at the peaks of the sawtooth.

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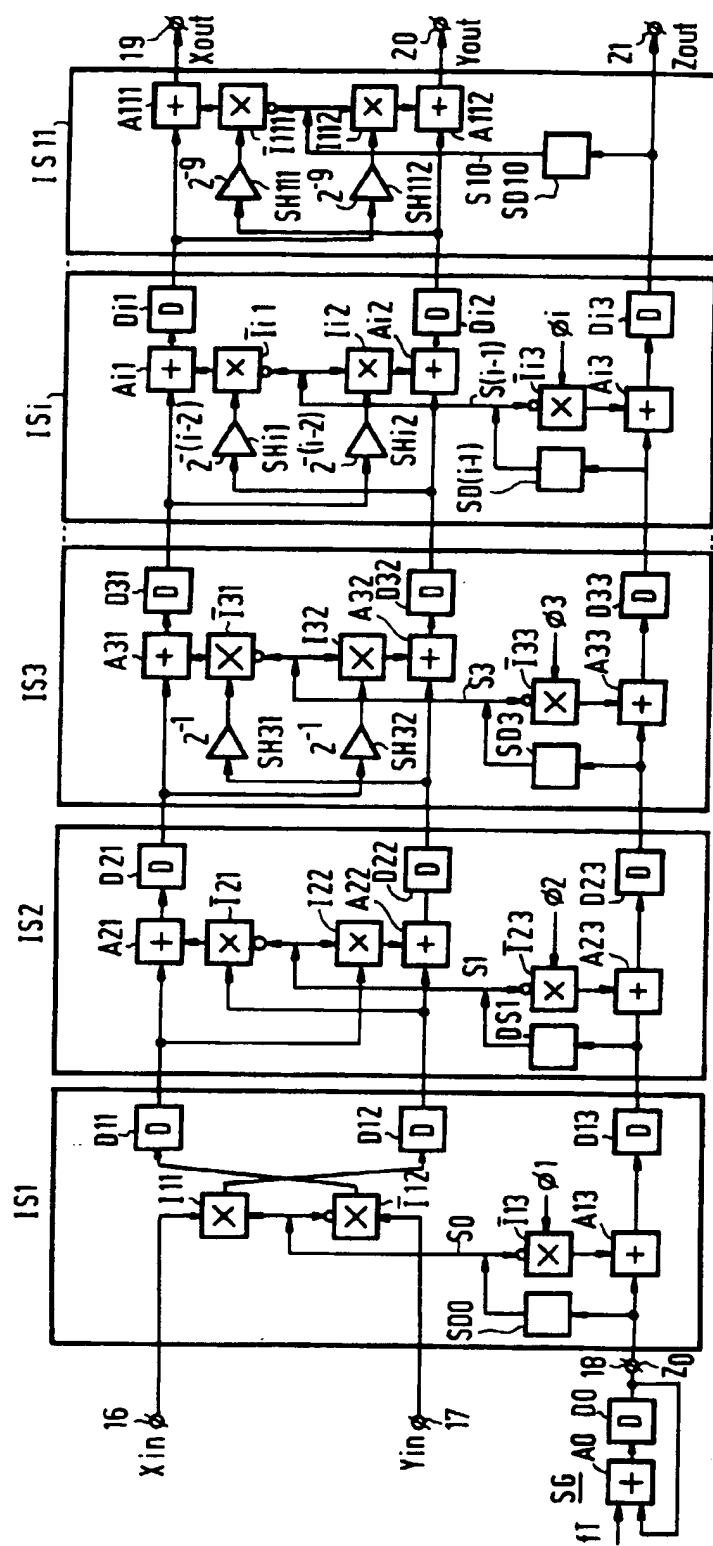


FIG.2

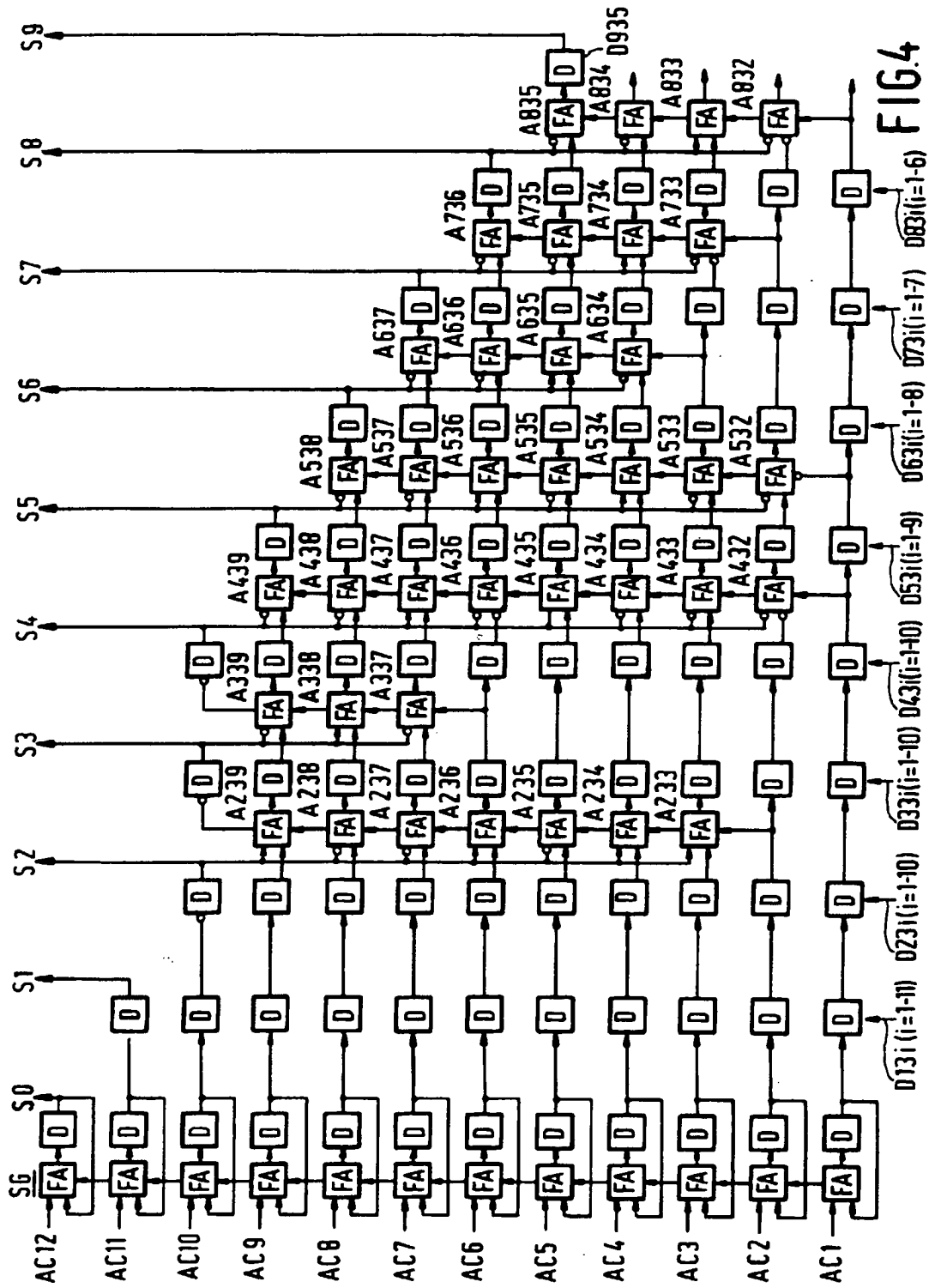


FIG. 4

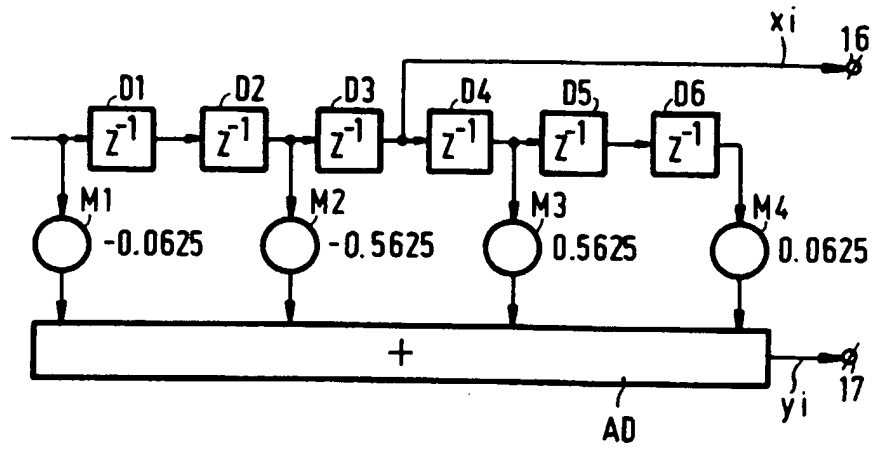


FIG. 5

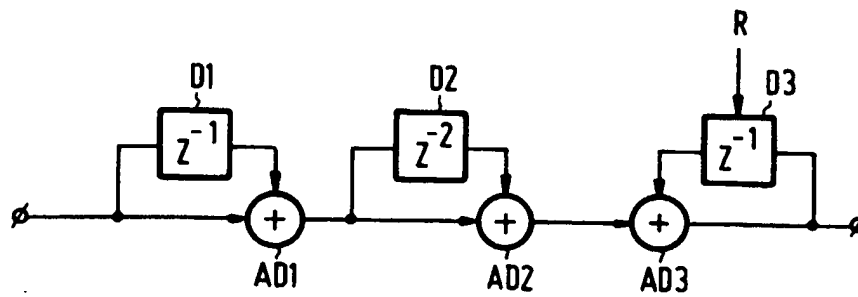


FIG. 6